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RADC-TR-88-307
Final Technical Report
November 1988

AD-A211 577

DIELECTRIC COVERED MICROSTRIP PATCH ANTENNAS

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Unclassified
SECURITY CLASSIFICATION OF THIS PAGE

REPORT DOCUMENTATION PAGE

1a. REPORT SECURITY CLASSIFICATION Unclassified	1b. RESTRICTIVE MARKINGS N/A
2a. SECURITY CLASSIFICATION AUTHORITY N/A	3. DISTRIBUTION/AVAILABILITY OF REPORT Approved for Public Release; Distribution Unlimited.
2b. DECLASSIFICATION / DOWNGRADING SCHEDULE N/A	
4. PERFORMING ORGANIZATION REPORT NUMBER(S) RADC-TR-88-307	5. MONITORING ORGANIZATION REPORT NUMBER(S)
6a. NAME OF PERFORMING ORGANIZATION Rome Air Development Center	6b. OFFICE SYMBOL (If applicable) EEAA
7a. NAME OF MONITORING ORGANIZATION	
6c. ADDRESS (City, State, and ZIP Code) Hanscom AFB Massachusetts 01731 5000	7b. ADDRESS (City, State, and ZIP Code)
8a. NAME OF FUNDING/SPONSORING ORGANIZATION Rome Air Development Center	8b. OFFICE SYMBOL (If applicable) EEAA
9. PROCUREMENT INSTRUMENT IDENTIFICATION NUMBER	
8c. ADDRESS (City, State, and ZIP Code) Hanscom AFB Massachusetts 01731-5000	10. SOURCE OF FUNDING NUMBERS

PROGRAM ELEMENT NO.	PROJECT NO.	TASK NO.	WORK UNIT ACCESSION NO.
62702F	4600	14	02

11. TITLE (Include Security Classification)

Dielectric Covered Microstrip Patch Antennas

12. PERSONAL AUTHOR(S)
Lisa M. Sharpe

13a. TYPE OF REPORT	13b. TIME COVERED	14. DATE OF REPORT (Year, Month, Day)	15. PAGE COUNT
In-House	FROM Sep 87 TO Sep 88	1988 November	24

16. SUPPLEMENTARY NOTATION

17. COSATI CODES		
FIELD	GROUP	SUB-GROUP
09	01	

18. SUBJECT TERMS (Continue on reverse if necessary and identify by block number)

Microstrip antenna, dielectric superstrate

19. ABSTRACT (Continue on reverse if necessary and identify by block number)

This in-house report provides detail on design and measurements of microstrip patch antennas with a dielectric cover layer.

20. DISTRIBUTION/AVAILABILITY OF ABSTRACT ☒ UNCLASSIFIED/UNLIMITED ☐ SAME AS RPT. ☐ DTIC USERS	21. ABSTRACT SECURITY CLASSIFICATION Unclassified
22a. NAME OF RESPONSIBLE INDIVIDUAL Lisa M. Sharpe	22b. TELEPHONE (Include Area Code) 617-377-4664 — 22c. OFFICE SYMBOL RADC/EEAA

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1. INTRODUCTION

Microstrip antennas have many properties that make them suitable for airborne and satellite communications systems. These antennas are low in cost and lightweight. For these reasons, Rome Air Development Center (RADC) is interested in verifying and augmenting existing design models for these antennas. This report will present the theory and results for modeling microstrip antennas that are covered with a sheet of dielectric material, as shown in Figure 1.

Bahl and Stuchly[1] give theoretical calculations for this configuration. By employing these calculations to design, build and test several microstrip antennas at various frequencies, we have attempted to verify this theory. We intend to use these design techniques as part of a software library that includes many necessary tools for building microstrip antennas.

There are several reasons for designing a microstrip antenna covered with a dielectric material. This configuration would allow the modeling of antennas with an integrated radome. A cover layer could possibly be used to support a polarizer; to mount additional antenna elements on top of the cover layer to provide bandwidth enhancements; or to be used as a dual frequency antenna.

2. TRANSMISSION LINE MODEL

The transmission line model for a covered patch is identical to that for the conventional patch with a single substrate. The superstrate is accounted for by modified parameters.

(Received for publication 13 December 1988)

1 Bahl, I.J., and Stuchly, S.S. (1980) Analysis of A Microstrip Covered with a Lossy Dielectric, *IEEE Trans. on Microwave Theory and Tech.*, **MTT-28** (No. 2):104-109.

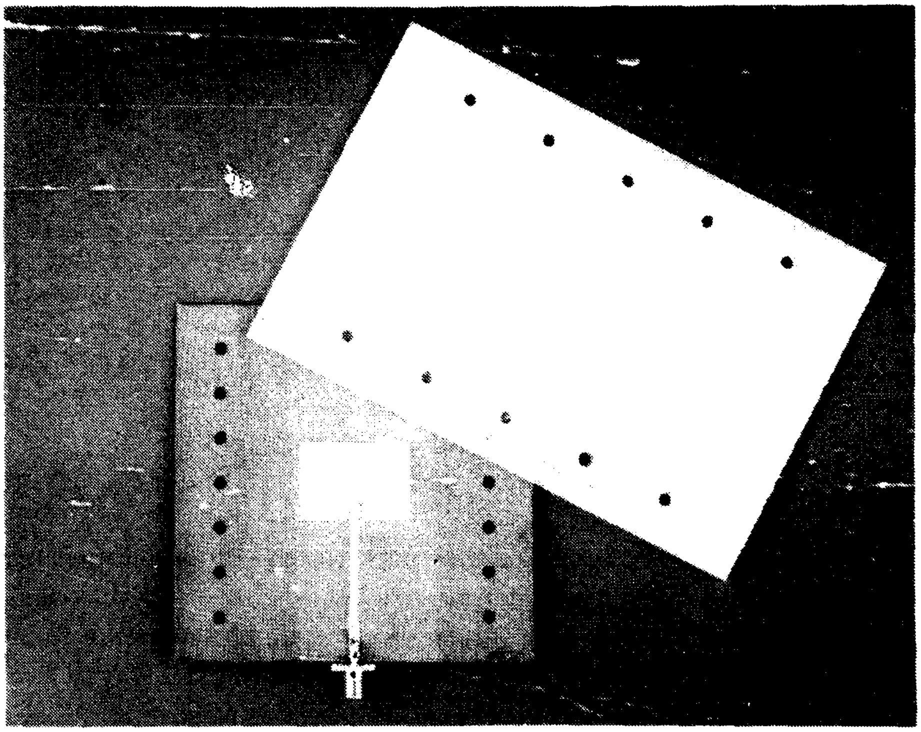

Figure 1. Rectangular Patch Antenna with Dielectric Cover Layer

Figure 2 shows a rectangular microstrip patch of width a, length b, and thickness t. It is on top of a sheet of dielectric material of thickness h_2 and is to covered by another dielectric sheet of thickness h_1. In the transmission line model, the radiating edges of the antenna are modeled as two slots slightly offset from the physical location of the edges of the patch. These slots are assumed to be of length a and width h_2. The admittance of these slots is found by [2]

$$Y = \frac{\pi a}{\lambda \eta_0}\left(1 - \frac{k^2 h_2^2}{24}\right) + j\,\frac{a}{\lambda \eta_0}\left[3.135 - 2\ln\left(kh_2\right)\right] \tag{1}$$

where

η_0 = intrinsic impedance of free space.

λ = wavelength

k = wave number

[2] Mullinix, D.A., and McGrath, D.T. (1986) *Rectangular Microstrip Patch Antenna Arrays*, RADC-TR-86-151, ADA179170.

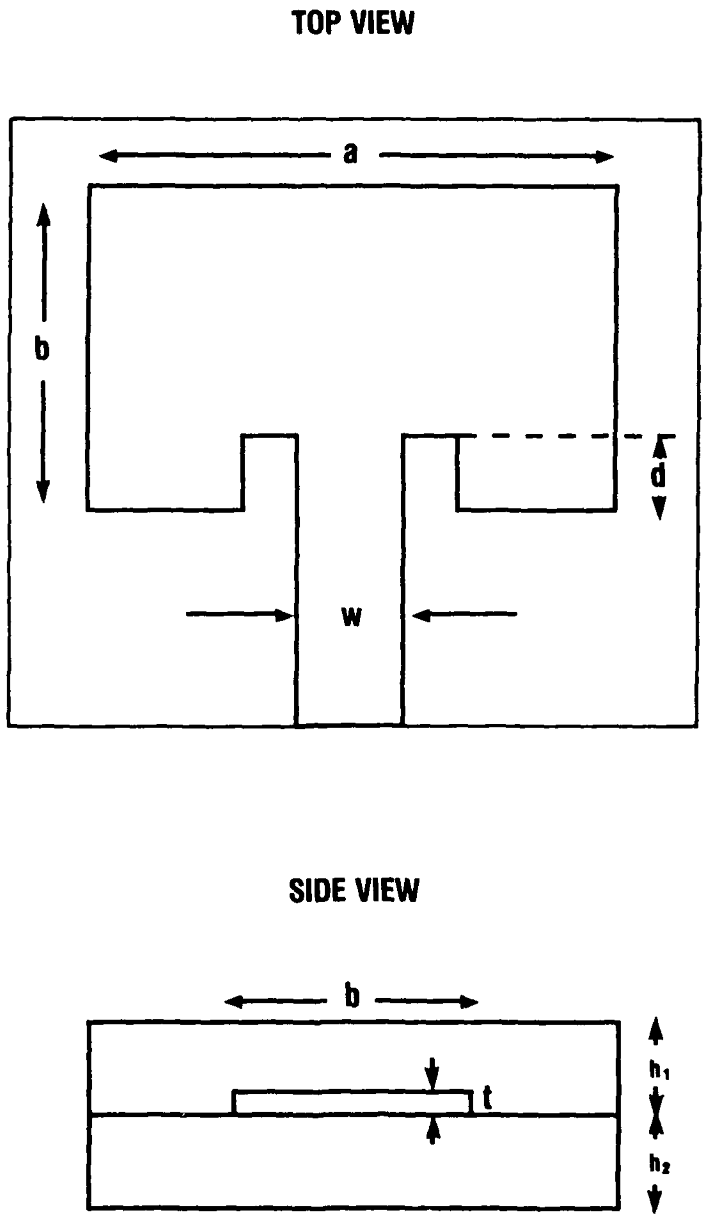

Figure 2. Dimensions of Rectangular Patch and Cover

The distance between the slots is extended by an amount

$$\Delta l = 0.412h_2\left(\frac{\varepsilon_{eff} + 0.3}{\varepsilon_{eff} - 0.258}\right)\left(\frac{a/h_2 + 0.264}{a/h_2 + 0.80}\right) \tag{2}$$

where ε_{eff} is the frequency compensated effective dielectric constant.

The frequency dependent dielectric constant ε_{eff} is calculated using the following equations[3].

$$\varepsilon_{eff} = \varepsilon_r - \frac{\varepsilon_r - \varepsilon_{eo}}{1 + G\left(f/f_p\right)^2} \tag{3a}$$

$$f_p = 15.66\, Zo/h_2 \tag{3b}$$

$$G = \sqrt{\frac{Z_o - 5}{60}} + .004\, Z_0 \tag{3c}$$

where Z_0 is the impedance corresponding to the patch width, and ε_r and h_2 are the dielectric constant and height, respectively of the substrate. ε_{ro} is the effective dielectric constant of the covered patch, as defined below.

Figure 3 shows an example of the frequency dependent dielectric constant. The results are for $\varepsilon_{r1} = 9$, $\varepsilon_{r2} = 2.54$, $h_1 = 0.125$, $h_2 = 0.0625$, $z = 100$ ohms. The change becomes more significant as the frequency is increased and therefore must be included in the calculations.

The modeled slots have a width equivalent to the height of the substrate layer, while the lengths are modeled as a and (a-w). The difference in length is due to the feed line partially blocking one of the radiating slots.

The calculation of the input admittance of the patch with a cover layer is then carried out using the following set of equations[1]

$$Z_0 = Z/\sqrt{\varepsilon_{eo}} \tag{4}$$

$$Z = \frac{1}{cC_0}$$

$$\varepsilon_{eo} = C/C_0$$

where Z_0 is the characteristic impedance of the microstrip line, C_0 is the capacitance of the transmission line structure without a cover layer, C is the capacitance with the cover, and ε_{eo} is the effective dielectric constant of the covered microstrip structure.

To calculate the capacitance of the covered microstrip, the equations given by Bahl and Stuchly[1] are used.

$$\frac{1}{C} = \frac{1}{\pi\varepsilon_0 Q^2}\int_0^{\infty} \frac{[f(\beta)]^2\, d(\beta h_2)}{\left[\varepsilon_{r1}\left(\dfrac{\varepsilon_{r1}\tanh(\beta h_1)+1}{\varepsilon_{r1}+\tanh(\beta h_1)}\right) + \varepsilon_{r2}\coth(\beta h_2)\right](\beta h_2)} \tag{5}$$

3 Bahl, I.J., and Bhartia, P. (1980) *Microstrip Antennas*, Artech House, Dedham, Mass.

$$f(\beta) = \int_{-\infty}^{\infty} f(x)dx$$

$$Q = \int_{-\infty}^{\infty} f(x)e^{j\beta x}dx$$

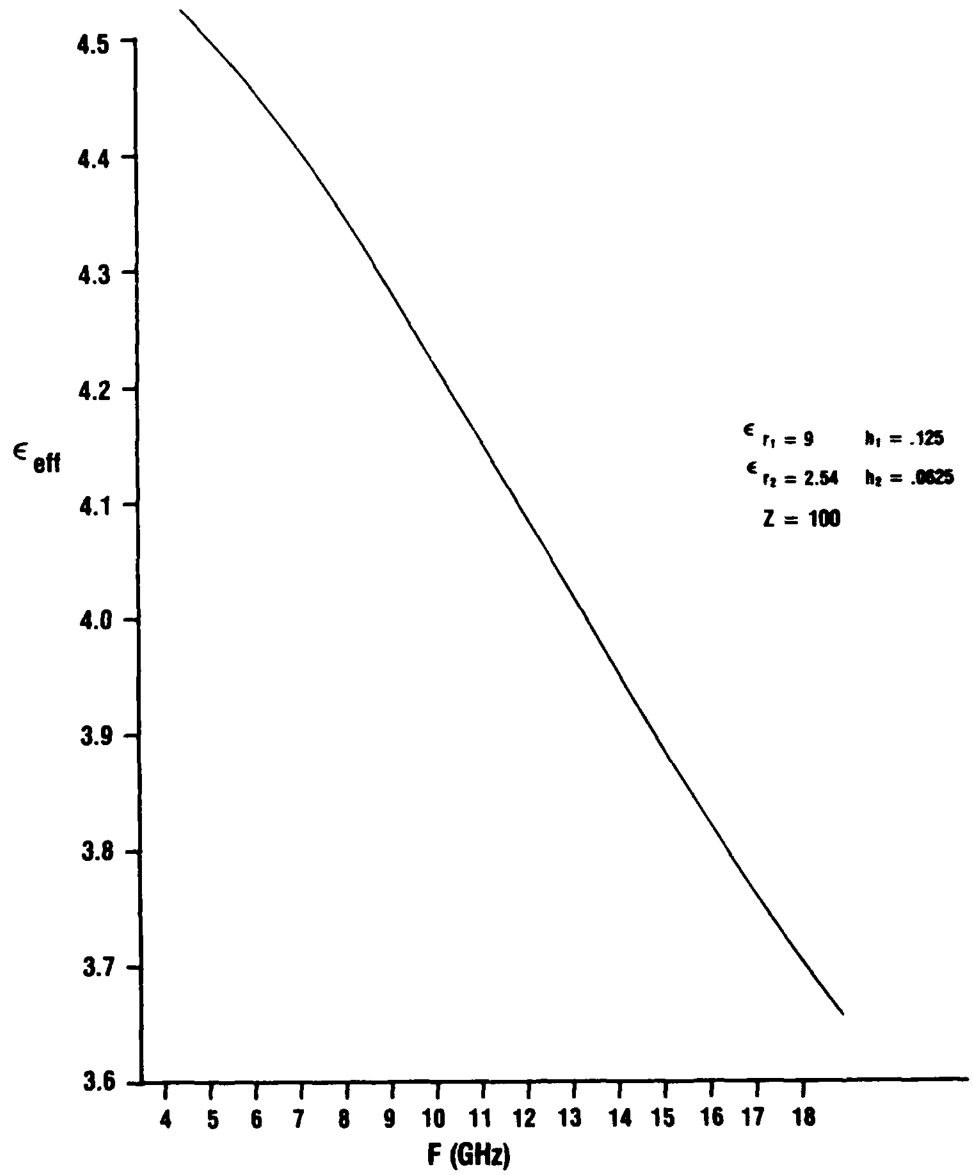

Figure 3. Dielectric Constant vs Frequency

By letting $\varepsilon_{r1} = 1$ or $h_1 = 0$ or both, Equation 5 can be used to find the capacitance of an uncovered microstrip transmission line.

The function f(x) used in the above formula is defined as the charge distribution on the strip conductor of width w. Bahl and Stuchly[1] assume this function to be

$$f(x) = \begin{cases} 1 + \left|\frac{2x}{w}\right|^3 & -w/2 < x < w/2 \\ 0 & \text{elsewhere.} \end{cases} \tag{6}$$

Our experiments showed that using a much simpler function for f(x) yielded identical results. Table 1 shows calculated center frequencies and dielectric constant for several patches of given dimensions using both f(x) given above and also

$$f(x) = \begin{cases} 1 + |x| & -w/2 < x < w/2 \\ 0 & \text{elsewhere.} \end{cases} \tag{7}$$

Using the simpler function for f(x) greatly simplifies the function $f(\beta)/Q$ without any significant change in the calculated results. This makes the calculation of the integral much quicker, using less computer time.

The theoretical calculations presented by Bahl and Stuchly[1] assume an infinitely thin strip. To compensate for the finite thickness of the strip, an additional multiplication factor must be included in the integral calculation, because the potential function $\phi(\beta, y)$ in the region above $y = h$ has an exponential behavior, $e^{-|\beta| y}$. This factor is included in the integral formula given previously. In our experiments, the thickness t, of the strip was t = 0.0014", since our substrates are typically clad with 1 oz/ft^2 copper. The additional multiplication factor is [4]

$$F(t) = \frac{1 + e^{-\beta t}}{2} \tag{8}$$

This term has the most significant effect on the results when w/h_2 is large.

A comparison was made between the effective dielectric constant, ε_{eo}, calculated using this quasi-static method and a full wave analysis.[5] Figure 4 shows the results of the two methods agree quite closely.

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[4] Yamashita, E., and Mittra, R. (1968) Variational Method for the Analysis of Microstrip Lines, *IEEE Trans. on Microwave Theory and Tech.*, **MTT-16** (No. 4):251-256.

[5] Herd, J.S. (1987) *Integral Equation Formulation of an Infinite Array of Electromagnetically Coupled Rectangular Microstrip Antennas*, A dissertation prospectus, University of Massachusetts.

Table 1. Effective Dielectric Constant for Different Current Distribution Functions

$\epsilon_{r1} = 1 \qquad h_1 = 0$

$\epsilon_{r2} = 4.4 \qquad h_2 = .0625$

$$f_1(x) = \begin{cases} 1 + \left| \dfrac{2X}{w} \right|^3 & -\dfrac{w}{2} < X < \dfrac{w}{2} \\ 0 & \text{elsewhere} \end{cases} \qquad f_2(x) = \begin{cases} |X| & -\dfrac{w}{2} < X < \dfrac{w}{2} \\ 0 & \text{elsewhere} \end{cases}$$

f_1(GHz) DESIGNED	f_1(GHz) CALCULATED	ϵ_{eff}	f_2(GHz) CALCULATED	ϵ_{eff}
4	4.04	3.939	4.01	3.776
5	4.99	3.886	4.97	3.718
6	5.91	3.840	5.94	3.673
7	6.85	3.779	6.88	3.636
8	7.79	3.761	7.82	3.606
9	8.75	3.726	8.77	3.581
10	9.63	3.694	9.65	3.558
11	10.59	3.664	10.60	3.539
12	11.39	3.636	11.40	3.520
13	12.38	3.609	12.39	3.504
14	13.14	3.585	13.15	3.488
15	14.13	3.562	14.13	3.473
16	14.97	3.540	14.97	3.459
17	15.86	3.519	15.86	3.445
18	16.79	3.500	16.79	3.432

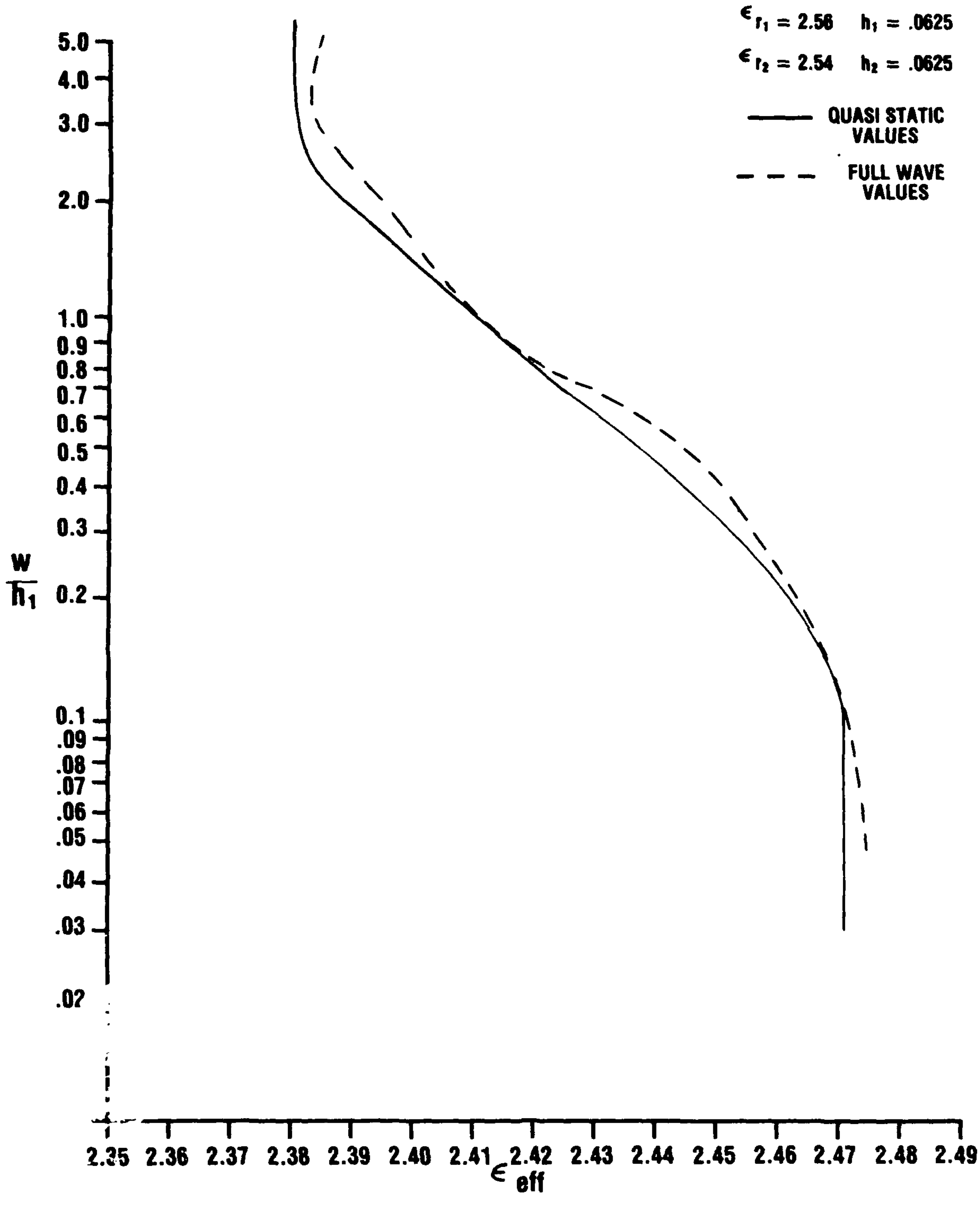

Figure 4. Calculated Effective Dielectric Constant Using Quasi-static Method and Full Wave Analysis

3. DESIGN OF PATCH RADIATOR

The calculations for the dimensions of a covered patch are carried out in a manner similar to that used for an uncovered microstrip patch. The resonant frequency of a patch is determined by the patch length b. The value of b that causes the patch input reactance to go to zero is the resonant length; it is determined iteratively from a starting value of $b = 0.49\lambda/\varepsilon_r$. The width of the patch is then calculated to be $a = 1.3$ b. Figure 5 shows the resonant length vs resonant frequency for $\varepsilon_{r1} = 9$, $\varepsilon_{r2} = 2.54$, $h_1 = 0.125$, $h_2 = 0.0625$. As can be seen from this graph, a higher resonant frequency does correspond to a smaller patch length.

Figure 6 shows the variation of the impedance as b is varied for a patch of resonant frequency 10 GHz. By stepping through successively larger values of b it is a simple matter to find the zero crossing of the reactance and therefore, the resonant patch length.

The resistance will reach its maximum value near the resonance length of the patch as shown in Figure 6. This is the resistance at the edge of the patch. The resistance decreases at points closer to the center of the patch, until directly in the center, the resistance is zero. Because the edge impedance is high, the feed line must be inset by a length d that corresponds to a matched impedance between the patch and the feed line.

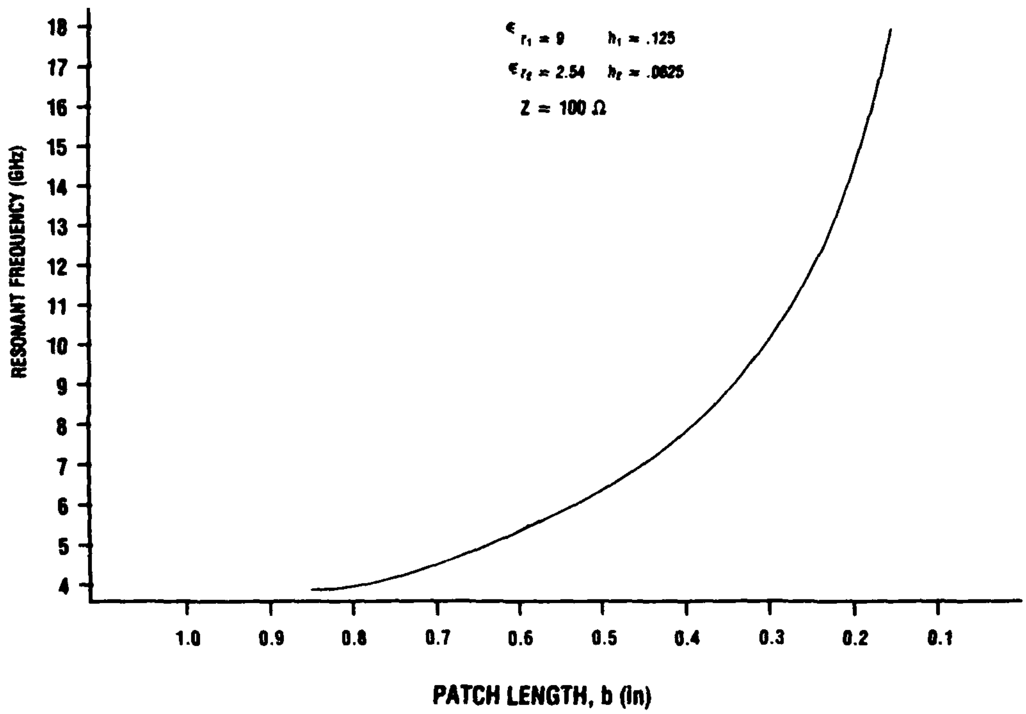

Figure 5. Resonant frequency vs Calculated Patch Length with Substrate $\varepsilon_{r2} = 2.54$ and Superstrate $\varepsilon_{r1} = 9$

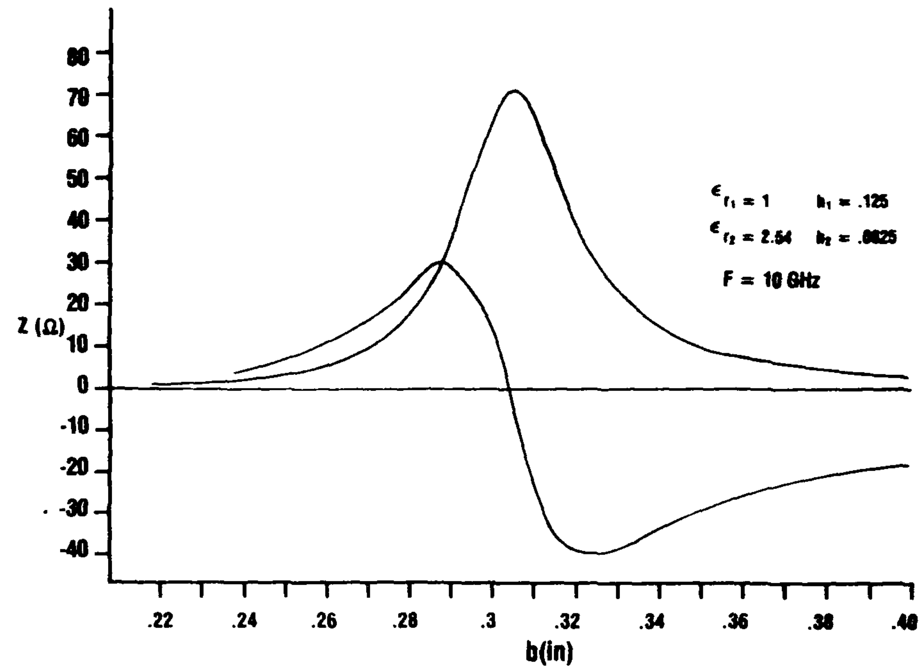

Figure 6. Impedance vs Patch Length for
10 GHz Patch with Dielectric Cover Layer

4. EXPERIMENTS AND RESULTS

The equations given by Bahl and Stuchly[1] to calculate the permittivity and the characteristic impedance of a covered microstrip transmission line of width w were evaluated numerically. A slight modification of this routine also allowed for the calculation of the width of the transmission line for a desired impedance.

Results of these routines are shown in Figures 7 and 8. Figure 7 shows the effective dielectric constant vs width of transmission lines for several different cover layers. The substrate used was ε_{r2} = 2.54, h_2 = 0.0625. The different cover layers were ε_{r1} = 1, h_1 = 0, ε_{r1} = 2.56, h_1 = 0.0625, and ε_{r1} = 9, h_1 = 0.125. The impedance was 50 ohms. These results were compared to those given by Bahl and Stuchly[1] and the agreement was found to be quite good. One of the cases run was for a relative dielectric constant of one, corresponding to no cover layer. Figure 7 shows the results for this case and also the results for the same calculation using well established equations for the modeling of microstrip line.[6] The agreement is very close for values of w greater than w = 0.05, while for smaller values of w, the agreement is not so close. This is because the numerical integration techniques used break down for very small values of w.

[6] Gupta, K.C., Gary, R. and Chadha, R. (1981), *Computer-Aided Design of Microwave Circuits*, Artech House, Dedham, Massachusetts.

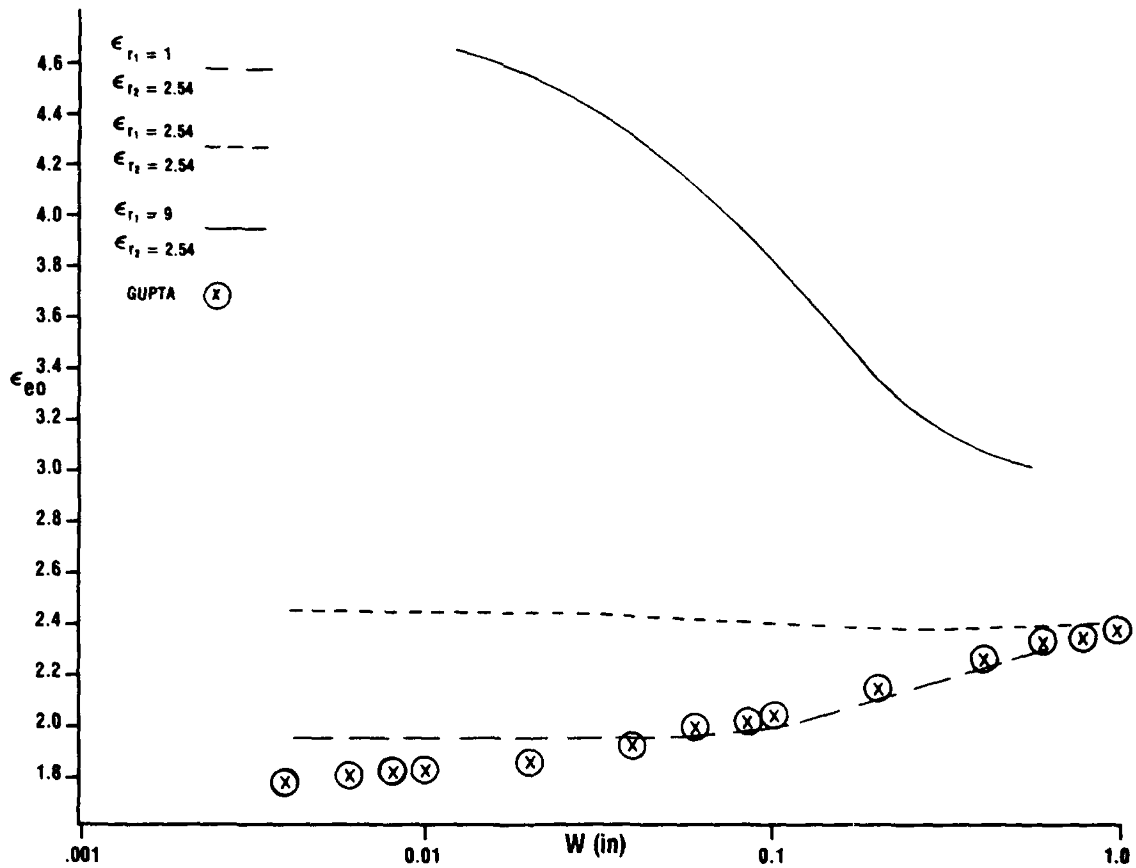

Figure 7. Width of Microstrip Line vs Dielectric Constant for Several Different Cover Layers

Figure 8 shows the results of the calculation of line width vs impedance for several different superstrates. Again, an uncovered patch was modeled and the results of this calculation were compared to the established results for uncovered microstrip. A very close agreement was once again seen.

Next, the theory of Bahl and Stuchly[1] was tested against measurements on actual patch radiators for the special case of no superstrate. The radiators were designed to resonate at specific frequencies between 4 and 18 GHz. The dielectric constants were $\varepsilon_{r2} = 4.4$ and $\varepsilon_{r1} = 1$. The height of the substrate was $h_2 = .0625$. It was found that the calculated resonant frequency did not give a very close approximation to the actual resonant. The discrepancies became more pronounced at the higher frequency. Figure 9 shows these results.

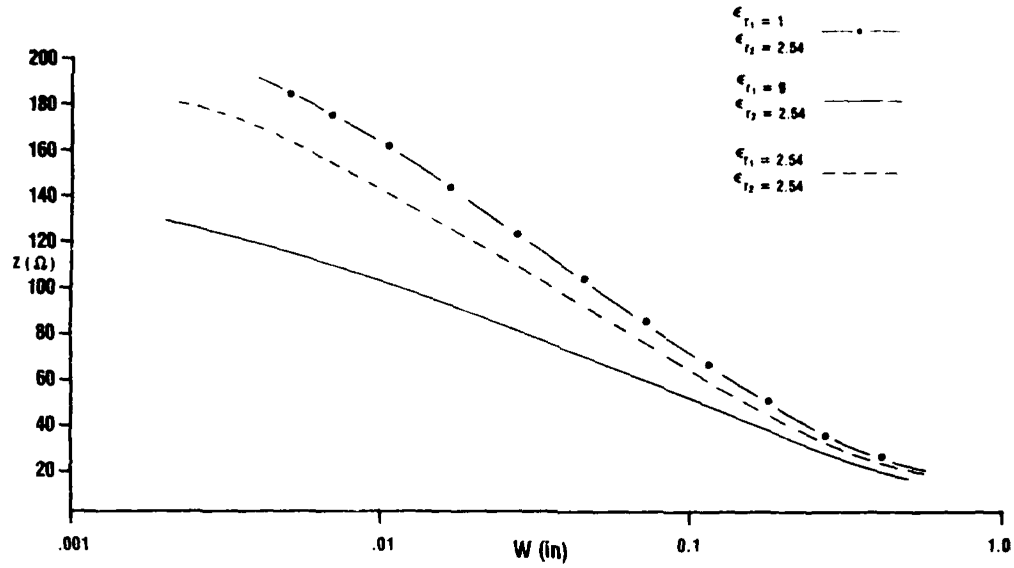

Figure 8. Width of Microstrip Line vs Impedance of the Line for Several Different Cover Layers

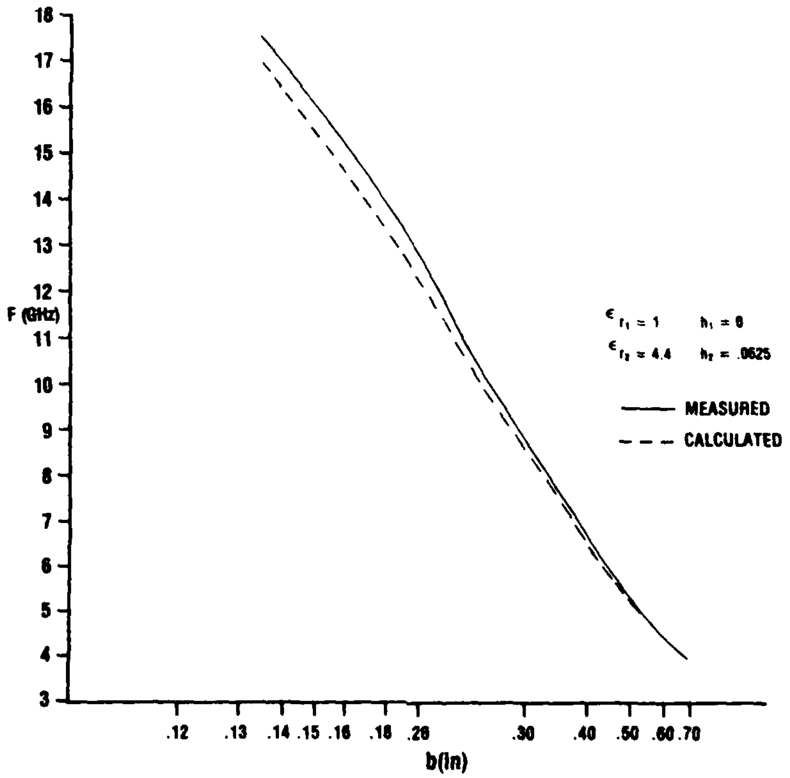

Figure 9. Resonant Frequency vs Patch Length for Measured Data and Calculated Data for Uncovered Patch Antennas

The next step was to calculate and measure the resonant frequency of these same patches when a cover layer was present. The cover layer used was $\varepsilon_{r1} = 2.56$, $h_1 = 0.0625$, thereby lowering the effective dielectric constant. Figure 10 shows the results. As can be seen, the measured and calculated resonant frequencies are very similar in this case.

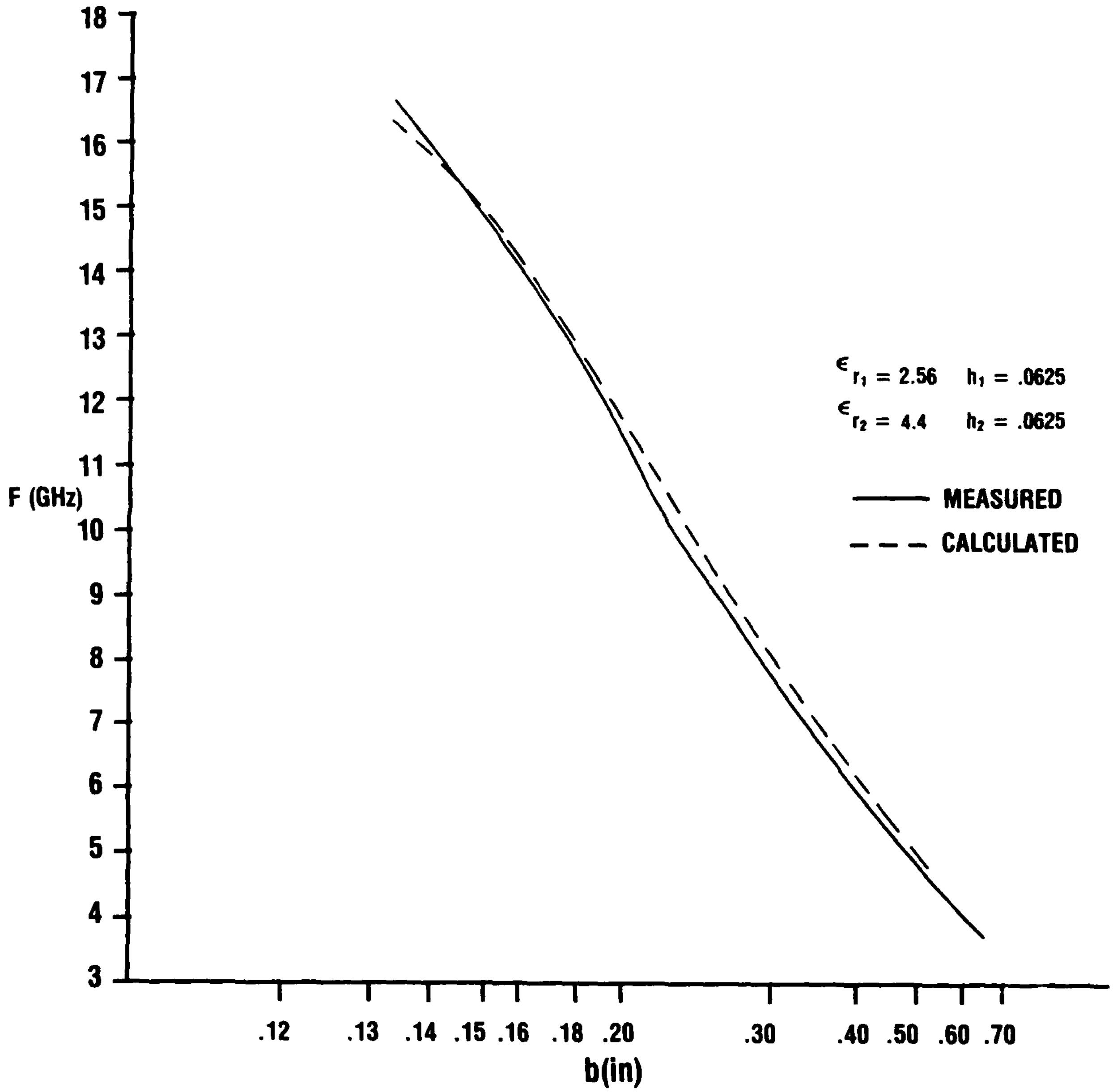

Figure 10. Resonant Frequency vs Patch Length for Measured Data and Calculated Data for Covered Patch Antenna

The routines were then used to design patches of frequencies 4-18 GHz on a substrate layer of ε_{r2} = 2.54, h_2 = 0.0625 and a cover layer with ε_{r1} = 9, h_1 = 0.125. The high dielectric constant of the cover layer was chosen to see if the theory would hold for a somewhat extreme case.

These patches showed poor resonances that were not at the designed frequencies. Figure 11 shows the best measured return loss for the 4 GHz covered patch As can be seen, the return loss is not very high and the resonant frequency is 3.87 GHz instead of 4.00 GHz.

Another set of patches was then made with ε_{r2} = 2.54, h_2 = 0.0625 and ε_{r1} = 2.56, h_1 = 0.0625. This set gave good results at the lower frequencies but as the frequency was increased the measured return loss was increasingly poor. At about 10 Ghz there was no longer a well-defined resonance. These results are shown in Table 2.

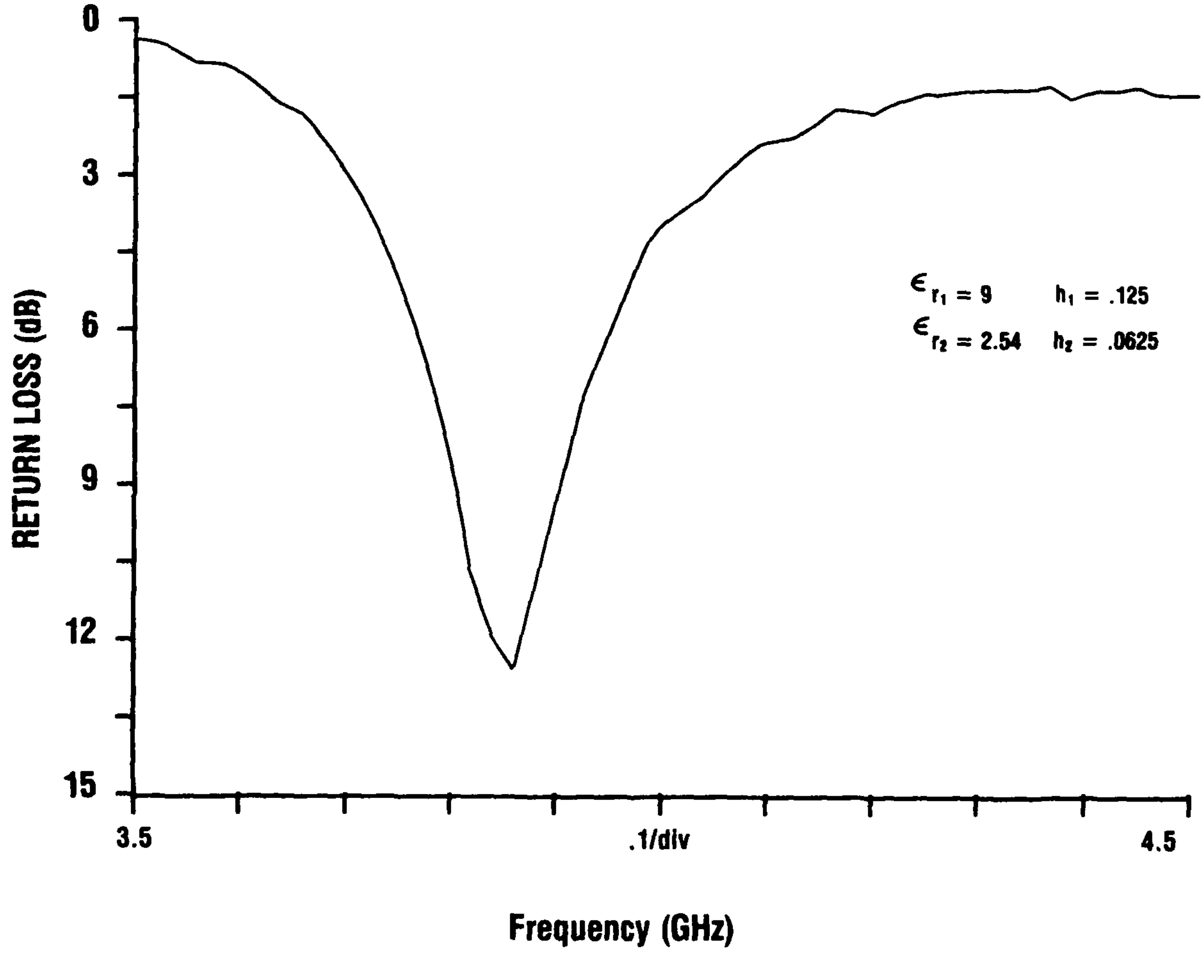

Figure 11. Measured Return Loss for 4 GHz Patch with Substrate ε_{r2} = 2.54 and Superstrate ε_{r1} = 9

Table 2. Parameters for Microstrip Patch Antennas with Substrate $\varepsilon_{r2} = 2.54$ and Superstrate $\varepsilon_{r1} = 9$

$\epsilon_{r1} = 2.56 \quad h_1 = .0625$

$\epsilon_{r2} = 2.54 \quad h_2 = .0625$

f(GHz) DESIGNED	f(GHz) MEASURED	RETURN LOSS (db)	$\Delta f/f$ %	a	b	d	h_2/λ_g
4	3.96	17.3	1	1.203	.869	.214	.033
5	4.92	17.2	1.6	.963	.686	.168	.041
6	5.90	18.0	1.6	.802	.565	.138	.049
7	6.88	12.9	1.7	.688	.479	.117	.058
8	7.76	10.2	3.0	.602	.415	.101	.066
9	8.76	8.6	2.7	.535	.365	.089	.074
10	9.72	8.3	2.8	.481	.326	.080	.082

There are a few possible causes of this limited success. One reason is that the equation used to calculate the admittance of the radiating slots is only valid for substrate thicknesses of $\frac{h}{\lambda} < 0.1$ (Eq. 1). As can be seen from Table 2, this condition is violated in about one-half of all cases. For this reason, poor results were achieved in the impedance match measurements. This is also the reason why the previous case, with $\varepsilon_{r1} = 9$ gave such poor results. This condition is violated at even lower frequencies when the dielectric constant is higher.

The slot admittance calculated using these equations is used in the calculation of the admittance of the patch. This impedance is used to determine the necessary inset feed length to match the patch antenna to the feed line. Because these equations are not valid unless $\frac{h}{\lambda} < 0.1$, an impedance mismatch, such as measured, would be expected. For ratios $\frac{h}{\lambda} > 0.1$, an alternate expression for the slot admittance which involves elliptic integrals should be used.[7]

[7] Harrington, R.F. (1961) *Time-Harmonic Electromagnetic Fields*, McGraw-Hill.

$$Y = Gs + jBs \tag{9}$$

$$\lambda\eta_o Gs = 2\int_0^{Kh/2} \frac{\sin^2 \omega d\omega}{\omega^2\sqrt{\left(\frac{Kh}{2}\right)^2 - \omega^2}} \tag{9a}$$

$$\lambda\eta_o Bs = 2\int_{\frac{Kh}{2}}^{\infty} \frac{\sin^2 \omega d\omega}{\omega^2\sqrt{\omega^2 - \left(\frac{Kh}{2}\right)^2}} \tag{9b}$$

These formulas will be used in a future study to see if improvements can be made.

The resonant frequency of a microstrip patch depends solely on the electrical distance between the two radiating edges. This electrical distance is $b\sqrt{\varepsilon_{eff}} + \Delta l$. The error in the resonant frequency, therefore, must be due to either an error in the length extension, Δl, given previously or in the calculated dielectric constant, ε_{eff}.

To determine if the source of the error was due to an incorrect dielectric constant, two transmission lines with different superstrates were fabricated and the frequency compensated effective dielectric constant of the transmission lines were measured. The two cases were both 50 ohm lines with a substrate dielectric constant of $\varepsilon_{r2} = 2.54$. For a cover layer of $\varepsilon_{r1} = 9$ the dielectric constant was calculated to be $\varepsilon_{eff} = 2.648$, while the measured value was $\varepsilon_{eff} = 2.64$. good agreement was also found for a cover layer of $\varepsilon_{r2} = 2.56$. In this case the measured value was $\varepsilon_{eff} = 2.43$ and the calculated value was $\varepsilon_{eff} = 2.479$. These results show that the approximation for the dielectric constant is accurate and therefore, the error in calculating the resonant frequency must be due to a poor value of Δl. The length extension derived by Hammerstad is for an uncovered patch, and therefore may not be accurate for the covered microstrip case. Future efforts will be made to determine the correct length extension.

5. CONCLUSIONS AND RECOMMENDATIONS

The mathematical equations given by Bahl and Stuchly[1] are valid for covered microstrip. The calculations of impedance vs line width and of effective dielectric constant vs line width showed results very close to well established models. The assumed current distribution on the microstrip transmission line does not seem to have a significant impact on the results. By using the simplest function for this distribution instead of the one given by Bahl and Stuchly[1], the same results can be achieved using less computations.

It appears that Bahl and Stuchly's[1] expressions give accurate parameters for the equivalent transmission line. However, when this equivalent line is used in the transmission line model for microstrip antennas, the accuracy is degraded. When a substrate of dielectric constant 2.54 was covered with a dielectric constant of 9, the effective dielectric constant was increased

significantly. Because of this, the approximations used were no longer valid. Future studies will focus on the changes necessary in the transmission line model to achieve better results for the covered microstrip patch antenna.

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